I-F AMPLIFIER FOR A RADAR RECEIVER

By Samuel A. Schwartz

I. INTRODUCTION

Radar receivers conventionally use a superheterodyne receiver with a diode mixer. The ability of the receiver to detect low-level echo signals from targets is limited primarily by the noise figure of both the mixer and the i-f amplifier.

The gain of the amplifier should be sufficient to raise the receiver's front-end generated noise and incoming signals of similar amplitude to a level which will cause linear operation of the second detector. Typically, the i-f amplifier should have a voltage gain of 120 db, which will raise a noise level of one microvolt to a detection level of one volt. The gain should be controllable over a range sufficient to allow proper operation on the reception of high-level echo signals. Additionally, it is desirable to control gain as a function of the time immediately following the transmitted pulse. This gain variation, commonly referred to as Sensitivity Time Control (stc), desensitizes the receiver for strong return signals immediately following the transmitted pulse. As time continues, the gain is increased to the maximum value determined by the setting of the Manual Gain Control (mgc), thus accommodating lower-level target signals.

A radar receiver was built employing the above design features for the complete solid-state-circuitry 25-mile X-band navigational radar set constructed by Fairchild Semiconductor. A high-reliability silicon planar transistor with a gain-bandwidth product of 900 mc, type 2N918, was used in all six stages. The design and performance of the i-f amplifier are described in this note; other application notes cover the video, sweep generator, modulator, etc.

The bandwidth of a radar i-f amplifier is determined by the transmitted pulse width, and is approximately twice the required video bandwidth. (See APP-76, "A Transistor Video Amplifier for Radar Display.") Thus, for this radar, which has an 0.4-μsec pulse width and a corresponding video bandwidth of 5 mc, the i-f bandwidth is 10 mc. Consistent with conventional radar i-f design practice, a 60-mc i-f center frequency is employed.

II. DESCRIPTION OF I-F AMPLIFIER

The complete i-f amplifier is composed of two separate amplifier sections, herein referred to as the preamplifier and the post amplifier. The terminations of the amplifiers are such that any length

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of interconnecting cable up to 100 feet has a minimum effect on the overall bandpass. The preamplifier is mounted in the antenna-transmitter assembly and the post amplifier in the radar indicator. Thus considerable physical separation between the two units may be obtained by means of the separable halves of the i-f amplifier with no degradation of bandpass performance. A block diagram of the i-f amplifier and mixer is shown in Fig. 1.

The i-f amplifier contains two sets of staggered triple amplifiers (triples) with an additional double-tuned circuit between the mixer and the first stage of the first triple.[1] The separation of the overall i-f amplifier into the preamplifier and the post amplifier is at the interconnection of the two triples. This arrangement has several important advantages:

1. Each triple can be tuned and neutralized for proper response separately from the other triple.
2. The gain of each triple is such that stability is easily obtained, whereas if a larger number of stages within the same physical location were used, considerably more effort would be required to produce the same degree of stability.
3. Since the stc and the mgc are connected to the second and third stages of the first triple, separating the preamplifier and post amplifier within the first triple would cause an undesirable shift in the bandpass. By sectionalizing the amplifier beyond the gain-controlled stages, proper bandshape is retained even with a long output cable leading to the post amplifier.
4. The bandpass response of the first triple can be aligned and tested independently without the post amplifier as the gain controls are varied.

The overall i-f amplifier gain is higher than the desired 120 db (voltage gain) to account for the matching loss of the mixer stage, the matching loss of the interconnecting cable between the amplifier halves, and the small second-detector loss. The preamplifier has a gain of 50 db, while the post amplifier has a higher gain of 75 db. Even though the two amplifier sections have the same number of stages, the gain of the preamplifier is lower because of the additional circuit damping required to maintain bandpass with the application of gain control. The gain may be reduced 40 db with practically no change in the bandpass.

Reverse agc was used rather than forward agc because the bandpass retention is superior with the application of agc voltage. Although the cross-modulation performance with reverse agc is not as good as with forward agc, there is small possibility that extraneous signals capable of producing cross-modulation effects would be present in the X-band.

The measured noise figure of the preamplifier is typically 4. 5 db using a device whose noise figure alone is 3. 5 db at 60 mc.

III. DESIGN OF I-F AMPLIFIER

The i-f amplifier, shown schematically in Fig. 2,* consists of six common-emitter stages that are matched and neutralized. The first three stages comprise the preamplifier and the last three stages, the post amplifier. Each group of three stages forms a staggered triple whose bandwidth is made greater than the overall i-f bandwidth to allow for shrinking when the triples are cascaded. The bandwidth of each triple for a pair of cascaded triples having an overall bandwidth of 10 mc is 10/0. 86, or 11. 6 mc.[2]

*See fold-out back page for Fig. 2.

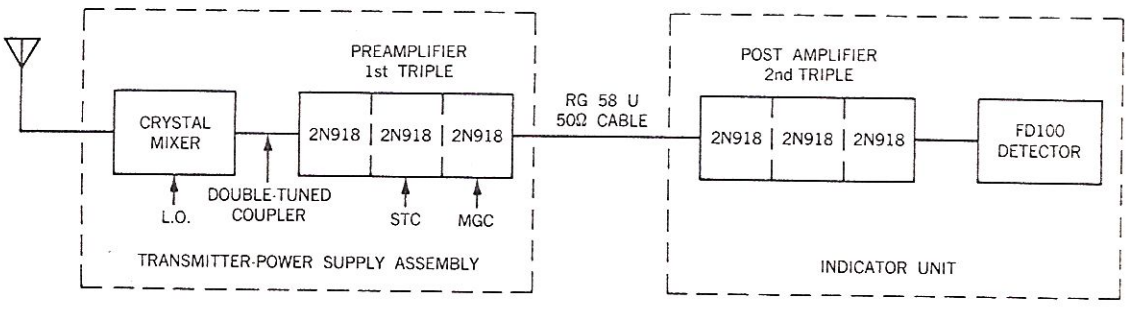


FIG. 1. Block diagram of radar receiver mixer and i-f amplifier.

The center frequency and bandwidth of the single-tuned interstage networks in each triple may be obtained from Table I.

TABLE I

TUNING CHART FOR FLAT STAGGERED TRIPLE

Stage	Frequency	Bandwidth (β)
1	f_0	β_t
2	$f_0 t + \beta_t$	$.097 (f_0 + 0.43 \beta_t)$
3	$f_0 t - \beta_t$	$.097 (f_0 - 0.43 \beta_t)$

The center frequency, f_0, of each triple is 60 mc with a bandwidth, β_t, of 11.6 mc. Substituting these values in the tuning chart yields the individual tuning frequencies and bandwidths shown in Table II. The bandpass characteristic of the second triple is shown in Fig. 3.

TABLE II

TUNING CHART FOR 60-MC TRIPLE

(β_t = 11.6 mc)

Stage	Frequency	Bandwidth
1	60 mc	11.6 mc
2	65 mc	6.3 mc
3	55 mc	5.34 mc

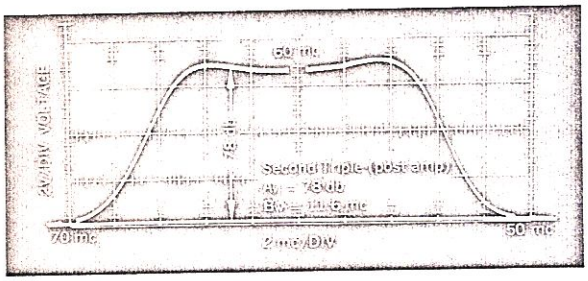


FIG. 3. Bandpass characteristic of second triple.

The correlation between the actual frequency of the interstage tuned circuits and the position of the circuit within the triple is theoretically arbitrary, but there are practical advantages to be gained by placing the band-center tuning elements at the output of the third stage of each triple. For example, the last i-f stage in the post amplifier, with the second detector load, can be tested independently of the other stages for proper bandwidth at the band-center frequency. The remaining two preceding i-f stages within the triple are then connected and the whole amplifier tuned for proper overall response. Tuning the last stage of the preamplifier to the center of the band (60 mc) results in the least shift of bandpass with the application of gain control when the cable to the post amplifier is long. Thus the line-up of the interstage tuned-circuit frequencies for the preamplifier and the post amplifier are: 65 mc, 55 mc, 60 mc, 65 mc, 55 mc, and 60 mc, consecutively.

The interstage tuned-circuit elements can be calculated after determining the shunt conductance and susceptance of the input and output ports of the transistors in adjacent stages. The following relation gives the ratio of total interstage conductance g_t to total susceptance b_t:

$$\frac{g_t}{b_t} = \frac{B_s}{f_s} \qquad (1)$$

$$\frac{g_1 + g_2 + g_3}{b_1 + b_2 + b_3} = \frac{B_s}{f_s} \qquad (2)$$

The conductance g_1 represents the input conductance of the driven stage transformed through the interstage transformation ratio n:1 to the output of the driving stage. The conductance g_2 represents the output conductance of the driving stage, while g_3 represents any additional conductance that may be required by the circuit, such as a shunt resistor. The susceptance b_1 represents the input susceptance of the driven stage transformed through the interstage transformation ratio to the output of the driving stage. The susceptance b_2 represents the output susceptance of the driving stage, and b_3 represents any additional susceptance which may be required by the circuit, such as a shunt capacitor.

An amplifier utilizing matched neutralized stages has a transformed input conductance g_1 of one stage equal to the output conductance g_2 of the

previous stage. Therefore Eq. (2) reduces to the following:

$$\frac{2g_1 + g_3}{b_1 + b_2 + b_3} = \frac{B_s}{f_s} \qquad (3)$$

In order to determine whether an additional shunt conductance g_3 or susceptance b_3 must be added, it is first necessary to calculate the ratio

$$\frac{2g_2}{b_1 + b_2}$$

for the transistors operated under proper bias conditions. If the ratio is smaller than B_s/f_s, an additional conductance g_3 must be added to the circuit, or if the ratio is larger, an additional susceptance b_3 must be added. A typical calculation will follow considering the interstage between the first and second stages of the post amplifier.

Fairchild transistors type 2N918 are operated under bias conditions providing maximum f_t of approximately 1 kmc. The corresponding collector voltage V_{CE} is 10 V and the collector current I_C is 5 mA. Under these conditions the input admittance is (3.0 + j3.3) millimhos and the output admittance is (0.25 + j0.9) millimhos. The test ratio

$$\frac{2g_2}{b_1 + b_2}$$

can then be calculated as follows:

$$\frac{2g_2}{b_1 + b_2} = \frac{2(0.25 \times 10^{-3})}{(3.3 + 0.9) \times 10^{-3}} = 0.119$$

$$\frac{B_s}{f_s} = \frac{6.5 \times 10^6}{65 \times 10^6} = 0.1$$

Since the ratio

$$\frac{2g_2}{b_1 + b_2}$$

is larger than B_s/f_s, an additional susceptance must be added to the circuit.* Therefore Eq. (3) reduces to the following:

$$\frac{2g_2}{b_1 + b_2 + b_3} = \frac{B_s}{f_s} \qquad (4)$$

From Eq. (4) the required additional susceptance b_3 is

$$b_3 = \frac{2g_2 f_s}{B_s} - (b_1 + b_2) \qquad (5)$$

The interstage transformation ratio n:1 can be calculated as follows for matched input and output conductances:

$$n = \sqrt{\frac{g_{11e}}{g_{22e}}} \qquad (6)$$

Substitution of numerical values yields

$$n = \sqrt{\frac{3 \times 10^{-3}}{0.25 \times 10^{-3}}} = 3.46$$

The input susceptance transformed to the primary of the coupling transformer becomes

$$b_1 = \frac{b_{11e}}{n^2} \qquad (7)$$

Substituting Eq. (6) in Eq. (7) yields

$$b_1 = \frac{b_{11e} g_{22e}}{g_{11e}} \qquad (8)$$

Equation (5) may thus be written in terms of transistor parameters.

*It is interesting to note the contrast between silicon planar transistors and vacuum tubes for use in broadband i-f amplifiers. The transistor exhibits a large conductance-to-susceptance ratio with a correspondingly large percentage bandwidth and therefore requires additional capacitance and hence susceptance to reduce the bandwidth to the required value. Just the opposite is true of tubes, which require the addition of damping resistance and thus shunt conductance to increase bandwidth to the desired value.

$$b_3 = \frac{2g_{22e}f_s}{B_s} - \left(\frac{b_{11e}g_{22e}}{g_{11e}} + b_{22e}\right) \qquad (9)$$

Substituting numerical values in Eq. (9) yields the necessary added shunt susceptance.

$$b_3 = \left[\frac{2(0.25 \times 10^{-3})(6.5 \times 10^{7})}{6.5 \times 10^{6}}\right]$$

$$- \left[\frac{(3.3 \times 10^{-3})(0.25 \times 10^{-3})}{3 \times 10^{-3}}\right]$$

$$+ 0.9 \times 10^{-3}$$

$$= 3.825 \text{ millimhos}$$

The required shunt capacitance C_p is

$$C_p = \frac{b_3}{2\pi f_s} = \frac{3.825 \times 10^{-3}}{2\pi(6.5 \times 10^{7})} = 9.4 \text{ pf} \qquad (10)$$

The total capacitive susceptance b_t is cancelled by the inductive susceptance of the interstage tuning inductance L_p. Therefore the value of L_p is chosen to resonate with the total capacitance C_t at the center frequency of the interstage coupling network.

$$L_p = \frac{1}{2\pi f_s b_t}$$

$$= \frac{1}{2\pi f_s (b_1 + b_2 + b_3)}$$

$$= \frac{1}{2\pi f_s \left(\frac{b_{11e}}{n^2} + b_{22e} + b_3\right)} \qquad (11)$$

Substitution of numerical values yields the required interstage inductance L_p.

$$L_p = \frac{1}{2\pi(65 \times 10^{6})\left[\frac{3.3 \times 10^{-3}}{(3.46)^2} + 0.9 \times 10^{-3} + 3.8 \times 10^{-3}\right]} = 0.491 \ \mu h$$

The transformation ratio between stages is obtained by winding additional turns on the inductor L_p. The turns ratio of the windings is then set by placing the appropriate number of turns on the additional winding.

The remainder of the interstage coupling networks in both the preamplifier and the post amplifier are calculated in a similar manner. (For the rest of the interstage network values, see the schematic, Fig. 2.) The network between the third stage of the first triple and the output cable takes into account the conductance of the resistive matching network terminated by the cable. All stages in the first triple have additional shunt conductance to reduce the percentage change in conductance at the interstage networks when gain control is applied.

The network between the third stage of the last triple and the second detector has an equivalent detector loading conductance of approximately twice the value of the detector load conductance. To allow a large output voltage from the second detector without limiting, the collector current of the last stage is 15 mA rather than 5 mA as used in other stages. The output admittance and input susceptance change only slightly when the collector current is increased 10 mA, but the input conductance changes appreciably. The interstage network design associated with this stage takes into account the parameters under the high-current condition.

For practical considerations, the prototype and final models of the i-f amplifier have circuit values slightly different from those calculated. Because of the small number of turns required on the primary winding, only a single turn is needed on the secondary, and therefore the transformers of the first and second stages in both the preamplifier and the post amplifier have a turns ratio of 3:1 rather

than 3.46:1 as calculated. The higher calculated ratio would necessitate somewhat less than one turn, which is hard to control in practice. Because of the lower-than-calculated turns ratio, the values of the primary tuning capacitors were slightly different from the calculated values. However, these changes caused no difficulty in obtaining proper bandshape.

Neutralization of each stage is accomplished by feedback from a phase-inverting winding on each interstage transformer to the base of the associated driving transistor through an adjustable neutralizing capacitor.

The overall stability of each triple is enhanced by the use of inductive coupling, rather than capacitive coupling, between stages and by arranging the input and output current loops so that the common coupling impedances are minimized. The simplified circuit of a single stage is shown in Fig. 4, which depicts only the signal current paths.

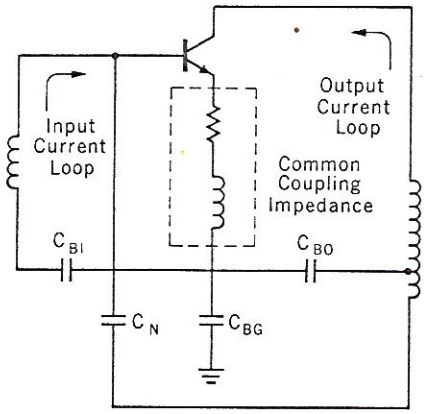


FIG. 4. Simplified schematic diagram of single amplifier stage.

Of particular importance is the fact that the input-loop and output-loop bypass capacitors C_{BI} and C_{BO}, respectively, are electrically and physically connected as closely as possible to the emitter of the transistor, or directly at the emitter pin of the transistor socket. Thus the common coupling impedance between the two loops is reduced to the resistive and inductive components of the lead wire and the bulk material of the transistor itself. The common-emitter connection joint is then electrically grounded by means of the emitter bypass capacitor C_{BG}. If the bypass capacitors C_{BI} and C_{BO} were tied directly to ground, either at a common point or separately, the series inductance of capacitor C_{BG} would increase the common coupling impedance and thus degrade circuit stability. Similar loops between stages are isolated by means of magnetic coupling between stages. I-f amplifier designs often take the neutralizing voltage from the secondary of the coupling transformer and feed it back to the base of the previous transistor. However, this arrangement again produces a common coupling impedance between stages. If the current through the neutralizing circuit is allowed to flow only within its associated stage, no interstage common coupling impedances are produced. Therefore it is preferable to add a neutralizing winding to the primary winding rather than to use a take-off point from the secondary winding.

The input network that matches the crystal mixer to the preamplifier consists of a broadband double tuned circuit. The bandwidth of this circuit is wide enough at 16 mc to prevent any overall bandwidth shrinkage when combined with the i-f amplifier. Fortunately, the input resistance of the first stage in the preamplifier is 330 Ω or approximately the same as the i-f output resistance of the crystal mixer. The double tuned circuit is transitionally coupled and has equal, and low, primary and secondary Q's. The design formulas[3] are as follows:

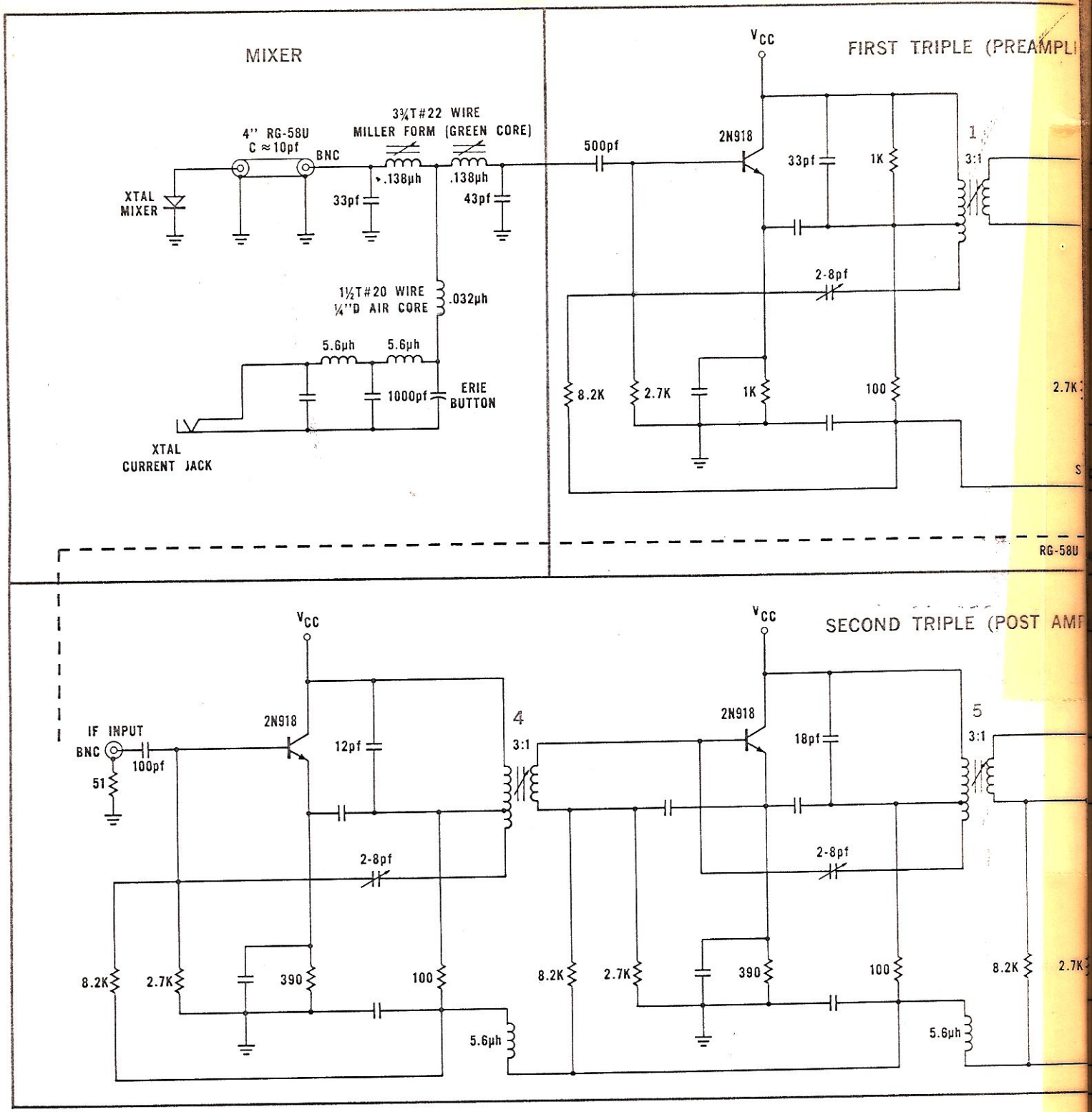
$$Q = \frac{\sqrt{2}\, f_o}{B} \tag{12}$$

$$KQ = 1 \tag{13}$$

$$f_1 = \frac{1}{2\pi\sqrt{L_1 C_1}} \tag{14}$$

$$f_2 = \frac{1}{2\pi\sqrt{L_2 C_2}} \tag{15}$$

FIG. 2. SCHEMATIC DIAG



C O I L D A T A

1
f_o = 65 mc
BW = 6.5 mc
L_p = .17 µh
P = 3 T
S = 1 T
N = 1¼ T
I_c = 2.5 mA

2
f_o = 55.4 mc
BW = 5.54 mc
L_p = .2 µh
P = 3 T
S = 1 T
N = 1¼ T

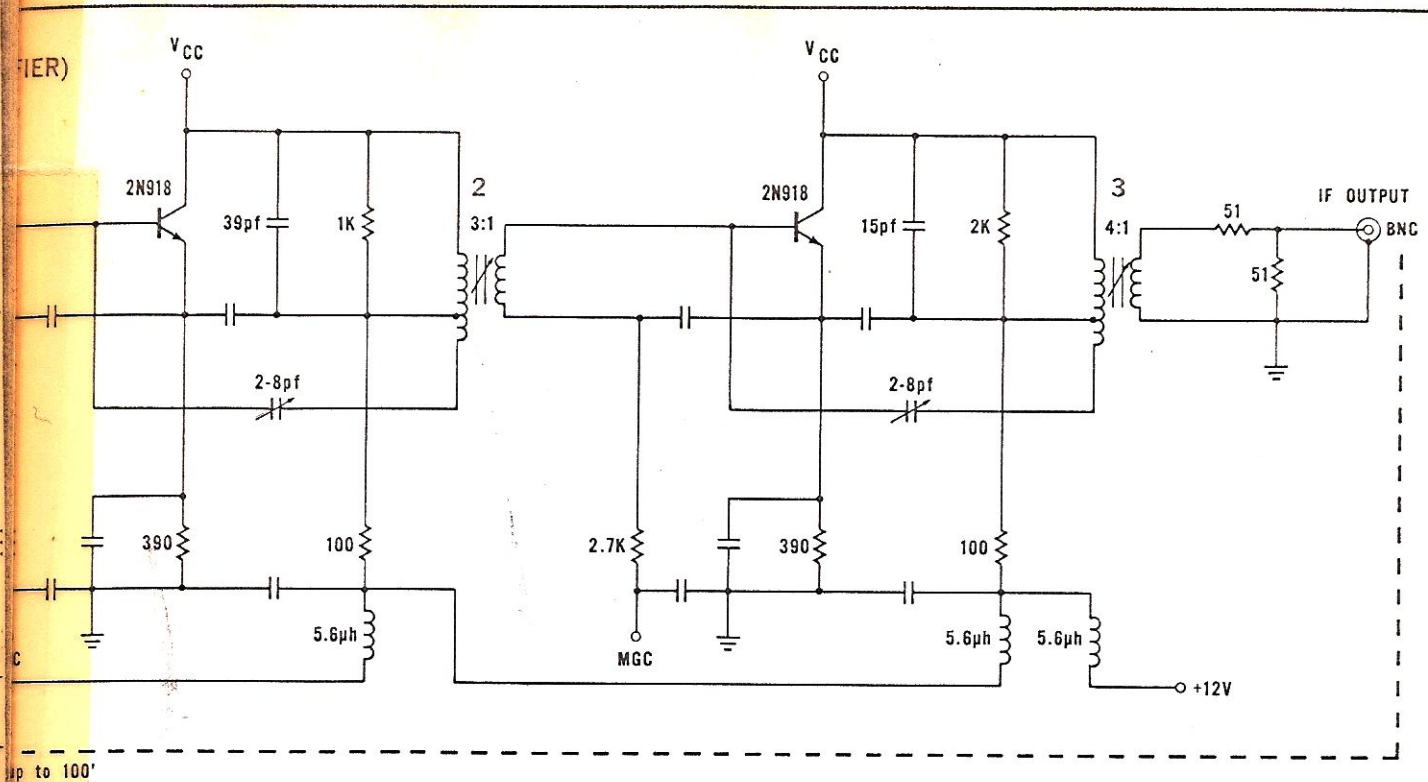
3
f_o = 60 mc
BW = 12 mc
L_p = .35 µh
P = 5 T
S = 1¼ T
N = 1¾ T

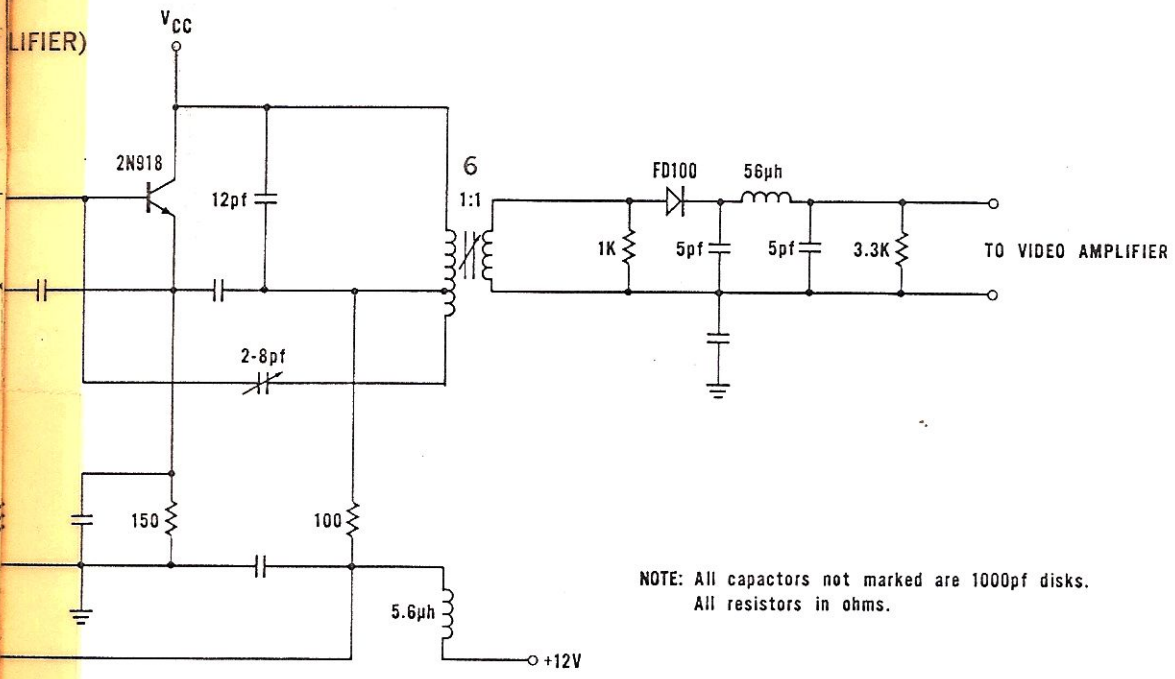
4
f_o = 65 mc
BW = 6.5 mc
L_p = .38 µh
P = 5 T
S = 1¾ T
N = 1¾ T
I_c = 6 mA

5
f_o = 55.4 mc
BW = 5.54 mc
L_p = .41 µh
P = 5 T
S = 1¾ T
N = 1¾ T
I_c = 6 mA

6
f_o = 60 mc
BW = 12 mc
L_p = .39 µh
P = 5 T
S = 5 T
N = 1¾ T
I_c = 15 mA

All coils not marked are bifilar wound on Sickles form, #30 enameled wire.





NOTE: All capactors not marked are 1000pf disks.
All resistors in ohms.

OVERALL DATA CHARACTERISTICS

FIRST TRIPLE (PREAMPLIFIER)

N.F.	≈ 6 db
f_o	= 60 mc
BW	= 12 mc
A_v	= 45 db
SADDLE	< 1 db

(Measured at 1 mV detected out detector referenced at 1 mV.)

SECOND TRIPLE (POST AMPLIFIER)

f_o	= 60 mc
BW	= 12 mc
A_v	= 78 db
V_{CC}	= 12 V

(Measured at 400 mV out referenced with TV Telonics Det. XD-3. Assumed both detectors same at 400 mV level.)

MIXER CHARACTERISTICS

GAIN db	VOLTAGE STC	VOLTAGE MGC
0	2.5	2.5
−5	1.29	1.2
−10	.98	.96
−15	.85	.85
−20	.77	.78
−25	.72	.7
−30	.68	.57

Substituting numerical values in Eqs. (12) and (13) gives the following:

$$Q = \frac{\sqrt{2}\,(60 \times 10^6)}{16 \times 10^6} = 5.3$$

$$5.3\,K = 1, \quad K = 0.189$$

For the coefficient of coupling K equal to 0.189 the following equations[4] hold:

$$\frac{f_1}{f_o} = \frac{f_2}{f_o} = 0.985 \qquad (16)$$

$$\frac{Q}{2\pi} = 0.845 \qquad (17)$$

The values of the primary and secondary tuning capacitors are as follows:

$$f_o R_1 C_1 = f_o R_2 C_2 = \frac{Q}{2\pi} \qquad (18)$$

Since $R_1 = R_2$

$$C_1 = C_2 = \frac{Q}{2\pi f_o R_2} = \frac{5.3}{2\pi\,(6 \times 10^7)\,(330)} \qquad (19)$$

$$= 42.6 \text{ pf}$$

(The cable that connects the preamplifier to the mixer crystal has a capacitance of 10 pf which is included in the primary tuning capacitance.)

The primary and secondary tuning frequencies being equal, are obtained as follows from Eq. (16):

$$f_1 = f_2 = 0.985\, f_o = 0.985\,(6 \times 10^7) = 59.1 \text{ mc}$$

The primary and secondary tuning inductors are obtained from Eqs. (14) and (15) as follows:

Since

$$f_1 = f_2 \text{ and } C_1 = C_2$$

then

$$L_1 = L_2 = \frac{1}{(4\pi)^2 (f_2)^2 C_2}$$

$$= \frac{1}{(4\pi)^2 \left(59.1 \times 10^6\right)^2 \left(42.6 \times 10^{-12}\right)}$$

$$= 0.17\ \mu\text{h}$$

The parallel tuned circuit is now converted to an equivalent π configuration[5] as shown in Fig. 5.

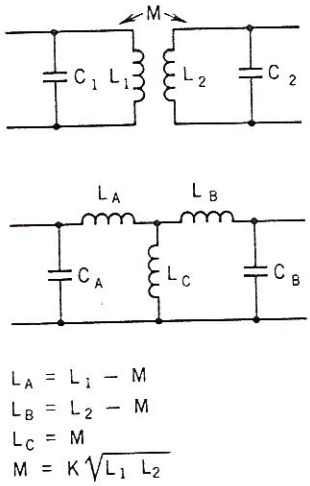


FIG. 5. Schematic diagram of double tuned circuit and T equivalent.

The mutual inductance M, which is equal to L_C, is as follows:

$$M = L_C = K\sqrt{L_1 L_2}$$

$$= 0.189\sqrt{(0.17 \times 10^{-6})\,(0.17 \times 10^{-6})}$$

$$= .0321\ \mu\text{h}$$

The series inductances L_A and L_B are as follows:

$$L_A = L_1 - M$$

$$= 0.17 \times 10^{-6} - 0.0321 \times 10^{-6}$$

$$= 0.138\ \mu\text{h}$$

Since $L_1 = L_2$

$$L_A = L_B = 0.138\ \mu\text{h}$$

The mutual coupling inductance L_C is a-c grounded by the input to the filter network that allows the measurement of crystal current.

LIST OF SYMBOLS NOT EXPLAINED ELSEWHERE

b_{11e}	Transistor input susceptance, common-emitter
B_s	Bandwidth of single tuned stage
b_{22e}	Transistor output susceptance, common-emitter
b_t	Total shunt susceptance
C_{BG}	Emitter bypass capacitor
C_{BI}	Input-loop bypass capacitor
C_{BO}	Output-loop bypass capacitor
C_1	Primary tuning capacitor
C_p	Total shunt capacitance
C_2	Secondary tuning capacitor
f_1	Primary tuning frequency
f_s	Center frequency of single tuned stage
f_2	Secondary tuning frequency
g_{11e}	Transistor input conductance, common-emitter
g_t	Total shunt conductance
g_{22e}	Transistor output conductance, common-emitter
L_A	Series inductor
L_B	Series inductor
L_C	Mutual coupling inductor
L_1	Primary inductor
L_p	Shunt interstage inductance
L_2	Secondary inductor

REFERENCE

1. Valley and Wallman, "Vacuum Tube Amplifiers", Radiation Laboratory Series, M. I. T., McGraw-Hill, 1949.

2. Ibid., Table 4.6, p. 191.

3. Ibid., pp. 217 - 219.

4. Ibid., Fig. 5.6 p. 219.

5. Thomas L. Martin, Jr., "Electronic Circuits," Prentice-Hall, Publishing Co., Englewood Cliffs, New Jersey, 1955.